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Low-complexity and modulation-format-independent carrier phase estimation scheme using linear approximation for elastic optical networks

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ABSTRACT

We propose a low-complexity and modulation-format-independent carrier phase estimation (CPE) scheme based on two-stage modified blind phase search (MBPS) with linear approximation to compensate the phase noise of arbitrary m-ary quadrature amplitude modulation (m-QAM) signals in elastic optical networks (EONs). Comprehensive numerical simulations are carried out in the case that the highest possible modulation format in EONs is 256-QAM. The simulation results not only verify its advantages of higher estimation accuracy and modulation-format independence, i.e., universality, but also demonstrate that the implementation complexity is significantly reduced by at least one-fourth in comparison with the traditional BPS scheme. In addition, the proposed scheme shows similar laser linewidth tolerance with the traditional BPS scheme. The slightly better OSNR performance of the scheme is also experimentally validated for PM-QPSK and PM-16QAM systems, respectively. The coexistent advantages of low-complexity and modulation-format-independence could make the proposed scheme an attractive candidate for flexible receiver-side DSP unit in EONs.

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1. Introduction

The rapid developments of digital signal processing (DSP) during the past decade have led to a global deployment of 100G/200G coherent systems. Moving forward to next generation 400G/1T systems, further reducing cost per bit and increasing network efficiency become the most important research in DSP techniques. Specifically, due to the dynamic, heterogeneous and unpredictable features of future internet traffic, the higher-order polarization multiplexing m-ary quadrature amplitude modulation (PM-mQAM) has been considered as a strong candidate to build up elastic optical networks (EONs) with flexible modulation formats (MFs), adaptive baud rate and variable transmission distances. Therefore, the DSP design targets will include higher spectral efficiency, improved tolerance to both linear and nonlinear noise, lower power consumption, higher flexibility and intelligence, and so forth [1–4]. In this case, to better support the "flexible" switch of MFs and line rate, the conventional solutions are employing alternative algorithms for different MFs, MF-recognition or MF-transparent algorithms [5]. However, MF-recognition or alternative algorithms are not beneficial to reduce the implementation complexity and lower the energy consumption as well as minimize the scale of DSP units [6]. Therefore, one of the pivotal components in EONs is the flexible DSP unit without dependence on MFs. In addition, as the modulation order increases, the phase information of the m-QAM signals dramatically impaired by the phase noise induced by the transmitter laser and the local oscillator (LO) due to the inherent shorter Euclidean distance, which results in a significant decrease of tolerance toward laser phase noise. Although narrow linewidth lasers could be utilized to maintain the performance, it causes high cost. Therefore, the deployment of CPE algorithm at receiver-side DSP units is most interested in the terms of the implementation cost and complexity [7,8]. As a consequence, another key design consideration for flexible DSP units is the low-complexity linewidth-tolerant CPE scheme, since it impacts on both the performance and the cost of the system.

To solve the two problems above, extensive studies of various complexity-reducing and/or modulation-format-independent CPE algorithms have been conducted [9–21]. For complexity-reducing CPE algorithms, the most popular examples among them are quadrature phase shift keying (QPSK) partitioning [14,15] and multi-stage blind

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phase search (BPS) [16–18]. QPSK partitioning schemes, with a lower computational complexity and easy implementing, are improved from the approach of classical Viterbi and Viterbi (V&V) [19] phase estimation. However, such algorithms, having a relatively low complexity but a stringent linewidth requirement, are not suitable for high order m-QAM coherent system to achieve complete discrimination for CPE due to the lack of sufficient dedicated symbols and specific amplitude. On the contrary, although the multi-stage BPS algorithms have very high linewidth tolerance and are applicable to a variety of MFs, they come with an expense of huge computational complexity as it still adopt fully blind search with fixed search step-size and require a considerable number of test phases which involves rotation, MF dependent decision, square and comparison in the complex plane. Moreover, the higher the modulation order is, the more the number of test phases is required to satisfy the estimation accuracy and the best linewidth tolerance. MF independent CPE algorithms have also been investigated. A universal phase-lock-loop based CPE was proposed in [20] but it comes with two weaknesses: one is that the feedback loop leads to a tradeoff between the linewidth tolerance and hardware speed, the other one is a convergence process which may cause wrong phase estimation. Based on a multistage phase error detection, a blind MF recognition method for CPE was reported [21]. However, this method only reduces the computational complexity of symbol decision process compared with the traditional BPS (hereinafter referred to as BPS) algorithm, but does not consider further reducing the number of test phases.

In this paper, we propose a low-complexity and modulation-format-independent CPE scheme based on two-stage modified BPS with linear approximation (LA-MBPS) to compensate the phase noise of arbitrary m-QAM signals in EONs. Firstly, the operating principle of LA-MBPS is elaborated in detail. What is more, the computation complexity, phase tracking accuracy and optimum parameter are investigated by numerical simulations. Compared with the traditional BPS scheme, the proposed LA-MBPS scheme has the advantages of lowering complexity, improving estimation accuracy and realizing independence to various MFs. Similar laser linewidth tolerance, in comparison with the traditional BPS scheme, is also observed form the simulation results. Finally, the slightly better OSNR performance of the scheme is experimentally demonstrated for PM-QPSK and PM-16QAM systems, respectively. The coexistent excellences make the proposed scheme a strong candidate for flexible receiver-side DSP unit in EONs.

2. Principle of LA-MBPS scheme

2.1. Approximate linear relationship between phase compensation error and corresponding error distance

It is well known that the BPS-based scheme can achieve nearly optimal estimation accuracy and linewidth tolerance for arbitrary m-QAM signals. However, these schemes have huge complexity since it requires a mass of test phases and a decision unit to provide the reference signal for Euclidean distance calculation. Thus, in the first place we focus on replacing the decision operations with simple taking absolute value operations.

Typically, quadrature imbalance, chromatic dispersion, timing error, polarization impairments and frequency offset are compensated before CPE, and the samples belong to the square or cross m-QAM constellations $(\pm a \pm j \cdot b)$, $a, b \in \{1, 3, 5, 7, \dots\}$, where m ($m = 2^n$, $n = 2, 3, 4, \dots$) is the highest level among all possible MFs in EONs.

The received symbol-rate sample before the CPE in a typical digital optical coherent receiver can be expressed as

$$r(k) = s(k)e^{(j\theta_k + j\phi)} + n(k)e^{j\varphi_k}, \tag{1}$$

where $s(k)e^{j\theta_k}$ denotes the kth transmitted symbol drawn from a QAM constellation, $n(k)e^{j\varphi_k}$ stands for additive complex white Gaussian noise, ϕ represents the phase noise induced by laser linewidth which remains unchanged within a proper time-window.

The received symbol-rate sample $r(k)$ is rotated by test phase angles ϕ_b,

$$Z(k, b) = r(k) \cdot e^{-j\phi_b}, \tag{2}$$

Then, we obtain the in-phase part $Z_I(k, b)$ and quadrature parts $Z_Q(k, b)$ of the rotated sample $Z(k, b)$, respectively. $Z_I(k, b)$ and $Z_Q(k, b)$ can be expressed as

$$\begin{aligned} Z_I(k, b) &= s(k)\cos(\theta_k + \phi - \varphi_b) + n(k)\cos(\varphi_k - \varphi_b), \\ Z_Q(k, b) &= s(k)\sin(\theta_k + \phi - \varphi_b) + n(k)\sin(\varphi_k - \varphi_b). \end{aligned} \tag{3}$$

As for the traditional BPS scheme, all rotated symbols are delivered into a decision circuit. The squared distance to the closest ideal constellation point is calculated at the complex plane. Different from the above method, we directly send the in-phase and quadrature parts of the rotated samples into a distance calculation module given by

$$\begin{aligned} &\begin{cases} Z_I(k, b, i) = abs(Z_I(k, b, i-1)) - 2^{\frac{n+\mathrm{mod}(n,2)}{2} - i}, \\ Z_Q(k, b, i) = abs(Z_Q(k, b, i-1)) - 2^{\frac{n+\mathrm{mod}(n,2)}{2} - i}. \end{cases} \quad i = 1, \dots, ceil(\frac{n}{2}) \\ &\begin{cases} d_I(k, b) = abs(Z_I(k, b, ceil(\frac{n}{2}))), \\ d_Q(k, b) = abs(Z_Q(k, b, ceil(\frac{n}{2}))). \end{cases} \end{aligned} \tag{4}$$

where mod$(n, 2)$ returns the modulus after division of n by 2, ceil$(n/2)$ rounds the elements of $n/2$ to the nearest integers greater than or equal to $n/2$, and abs (.) stands for taking absolute value. $Z_I(k, b, 0)$ and $Z_Q(k, b, 0)$ are the in-phase part $Z_I(k, b)$ and the quadrature parts $Z_Q(k, b)$ of the rotated sample $Z(k, b)$ respectively, as expressed in Eq. (3). $d_I(k, b)$ and $d_Q(k, b)$ represent the real and imaginary parts of the final error distance.

In order to further elaborate the principle of Eq. (4), as illustrated in Fig. 1, the signal constellation symmetrical folding processes are given by taking ideal 256-QAM as an example. From the figure we can see that the 256QAM signal is folded into the first quadrant by the first taking absolute value operation of the real and imaginary parts. Then it can be translated into a new 64QAM signal after the real and imaginary parts all minus 8. Next, the 64QAM signal will be folded into the first quadrant by the second taking absolute value operation and then translated into a new 16QAM signal after the real and imaginary parts all minus 4. We repeat the operations as above mentioned until it becomes QPSK. After the fourth taking absolute value operation, QPSK signals will be folded into the first quadrant as a point. Finally, we get the corresponding error distance by the real and imaginary parts of the point subtracting 1, respectively. It is worth noticing that the constellation folding processes of other lower modulation formats are also summarized in the figure. In addition, as not all 8QAM constellation points could overlap with other square QAM signals, it is noteworthy that the folding mechanism here can only be applied to rectangular or vertical 8QAM constellation with points of $(\pm a, \pm j \cdot b)$, $a, b \in \{1, 3\}$.

Obviously, $d_I(k, b)$ and $d_Q(k, b)$, the real and imaginary parts of the error distance between the rotated sample and the closest ideal constellation point can be represented as

$$\begin{aligned} d_I(k, b) &= s(k)\cos(\theta_k + \phi - \varphi_b) + n(k)\cos(\varphi_k - \varphi_b) - s(k)\cos(\theta_k), \\ d_Q(k, b) &= s(k)\sin(\theta_k + \phi - \varphi_b) + n(k)\sin(\varphi_k - \varphi_b) - s(k)\sin(\theta_k), \end{aligned} \tag{5}$$

In order to smooth out the influences of additive white Gaussian noise (AWGN) on error distances calculation, the error distances of L consecutive received symbols rotated by the same test phase angle φ_b

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